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6. AUTHOR(S) Xiaodong Yue and H. Howard Fan					
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Near-Far Resistance of Alamouti Space-Time Coded CDMA Communication Systems

Xiaodong Yue and H. Howard Fan
University of Cincinnati

Abstract

In this paper, the near-far resistance of the MMSE detector is derived for the code division multiple access (CDMA) communication systems with the Alamouti space-time block coding. It is shown that transmitter/receiver diversity has benefits on the near-far resistance performance while receiver diversity has more impact on the near-far resistance.

Keywords: Code division multiaccess, Near-far resistance, Spatial diversity

1 Introduction

New multimedia applications in the field of mobile communications require high data rate transmission under the frequency selective fading environment. One of the most promising techniques to achieve this goal is space-time (ST) coding. Using multiple transmit and receive antennas, space-time processing involves exploitation of spatial diversity and coding gain over an uncoded system without sacrificing the bandwidth. A simple space-time block coding (STBC) scheme developed by Alamouti [1] has been adopted in several wireless standards such as IS-136, wideband CDMA (W-CDMA) and CDMA-2000 [2]. It has been shown that this technique can significantly improve system performance. However, the well-known near-far problem in a multiuser setting still places fundamental limitations on the performance of ST coded CDMA communication systems. Therefore, near-far resistance remains one of the most important performance measures for ST coded CDMA systems. Furthermore, it would be interesting to find out if the transmitter/receiver diversity has benefits on near-far resistance. It is worthy to mention that near-far resistance also depends on the type of the receive detector. In this paper, the near-far resistance of the widely used MMSE detector is derived for the Alamouti ST coded CDMA communication systems. It is shown that transmitter/receiver diversity has benefits on the near-far resistance performance while receiver diversity has more impact on the near-far resistance.

2 Signal model

Consider an asynchronous CDMA system with J active users in a 2 transmit and 2 receive antennas configuration. In Alamouti's original scheme, each data block contains two symbols. Specifically, for each user j , two information symbols $b_{1,j}$ and $b_{2,j}$ are transmitted over two symbol intervals. At the first time interval, the symbol pair $(b_{1,j}, b_{2,j})$ is transmitted across the two transmitter antennas, and at the second time interval, the symbol pair $(-b_{2,j}^*, b_{1,j}^*)$ is transmitted across the two transmitter antennas, where the superscript "*" denotes the complex conjugate. Each user uses two different spreading codes with length L_c for $b_{1,j}$ and $b_{2,j}$ [3]. The channel model assumed in this paper is slowly varying frequency-selective multipath fading channel. Due to the lack of space, we directly give the

Corresponding author: H. Fan, Dept. of ECECS, ML 30, University of Cincinnati, Cincinnati, OH 45221 USA, phone: (513) 556-4765, e-mail: h.fan@uc.edu

expression for the discrete-time signal model of Alamouti ST coded CDMA communication system under frequency-selective multipath fading channels without derivation. Refer to [3] and references therein for a detailed derivation. Since the detector does not know the starting time of each desired user symbol, then similar to a chip level detector for an asynchronous CDMA system, a smoothing factor M is needed to capture a complete desired user symbol. By collecting the received signals at each receiver antenna into a vector, the total received signal vector $\mathcal{X}_M(n)$ observed in additive white Gaussian noise $\mathbf{v}(n)$ can be expressed as [3]

$$\mathcal{X}_M(n) = \mathcal{H}\mathbf{b}(n) + \mathbf{v}(n) \quad (1)$$

where n is the time index, \mathcal{H} is a block Toeplitz channel matrix, and $\mathbf{b}(n)$ is the transmitted information symbols. Since the channel effect on received energy can always be incorporated into a diagonal amplitude matrix \mathbf{A} , without loss of generality, we assume that the columns of the channel matrix \mathcal{H} are all *normalized* in the following sections. Then (1) can be rewritten as

$$\mathcal{X}_M(n) = \mathcal{H}\mathbf{A}\mathbf{b}(n) + \mathbf{v}(n) \quad (2)$$

3 Near-far resistance of MMSE receiver for Alamouti ST coded multiple antennas CDMA systems

The following assumptions will be made throughout this paper. **AS1** : The symbols $b_j(n)$ are uncorrelated in time, with variance 1 (since symbol energy can also be absorbed into the diagonal matrix \mathbf{A}) for each j where the subscript j denotes the j th user. $b_i(n)$ and $b_j(n)$, $i \neq j$, are also uncorrelated. **AS2** : The noise is zero mean white Gaussian, and uncorrelated across different receive antennas. **AS3** : The channel matrix \mathcal{H} is of full column rank (known as the identifiability condition in the blind multiuser detection/equalization literature). Note **AS3** is a reasonable assumption in practice considering the randomness of the multipath channels [4].

Without loss of generality, we assume that the d th symbol in $\mathbf{b}(n)$ is the desired transmitted symbol of the desired user and simply denote it by $b_d(n)$ ¹. Therefore, the MMSE detector weight vector is given by $\mathbf{w}_{mmse} = \mathbf{R}^{-1} \mathbf{H}_d$ [3][5], where \mathbf{R} is the autocorrelation matrix of the received signal $\mathcal{X}_M(n)$ and \mathbf{H}_d is the d th column in \mathcal{H} corresponding to the desired transmitted symbol $b_d(n)$. It is also well known that when noise approaches zero, the zero forcing (ZF) detector is proportional to the MMSE detector [6], $\mathbf{w}_{zf} = \alpha \mathbf{w}_{mmse}$, where α is a constant. Therefore, both detectors have the same near-far resistance [6]. We have the following Proposition on near-far resistance of Alamouti ST coded CDMA systems

Proposition 1 *The near-far resistance of the MMSE detector for Alamouti ST coded CDMA system (2) is $\bar{\eta}_d = \frac{1}{(\mathcal{H}^H \mathcal{H})_{(d,d)}^{-1}}$, where the subscript (d,d) denotes choosing the element at the d th row and the d th column.*

Proof: By applying the zero forcing detector to the received signal vector $\mathcal{X}_M(n)$, the output contains only the useful signal and ambient Gaussian noise. The amplitude of the useful signal at the output is $\mathbf{w}_{zf}^H \mathbf{H}_d A_d b_d(n)$ where A_d is the received signal amplitude of $b_d(n)$. Therefore, the energy of the useful signal at the output is $E_s = E [\mathbf{w}_{zf}^H \mathbf{H}_d A_d b_d(n) b_d^*(n) A_d \mathbf{H}_d^H \mathbf{w}_{zf}] = A_d^2 \mathbf{w}_{zf}^H \mathbf{H}_d \mathbf{H}_d^H \mathbf{w}_{zf}$. The variance of the noise is $E_n = \sigma^2 \mathbf{w}_{zf}^H \mathbf{w}_{zf}$ where σ^2 is the power spectral density of white Gaussian noise. Using the definition in [6], the asymptotic multiuser efficiency (AME) for the desired transmitted symbol is

$$\eta_d = \lim_{\sigma \rightarrow 0} \frac{\sigma^2 [Q^{-1}(P_d(\sigma))]^2}{A_d^2} = \lim_{\sigma \rightarrow 0} \frac{\sigma^2 E_s / E_n}{A_d^2} = \lim_{\sigma \rightarrow 0} \frac{\sigma^2 A_d^2 \mathbf{w}_{zf}^H \mathbf{H}_d \mathbf{H}_d^H \mathbf{w}_{zf}}{\sigma^2 \mathbf{w}_{zf}^H \mathbf{w}_{zf} A_d^2}$$

¹Unlike the notation in **AS1**, the subscript d of $b_d(n)$ only represents its position in $\mathbf{b}(n)$.

$$\begin{aligned}
&= \lim_{\sigma \rightarrow 0} \frac{\mathbf{w}_{mmse}^H \mathbf{H}_d \mathbf{H}_d^H \mathbf{w}_{mmse}}{\mathbf{w}_{mmse}^H \mathbf{w}_{mmse}} = \lim_{\sigma \rightarrow 0} \frac{\mathbf{H}_d^H (\mathcal{H} \mathbf{A}^2 \mathcal{H}^H + \sigma^2 \mathbf{I})^+ \mathbf{H}_d \mathbf{H}_d^H (\mathcal{H} \mathbf{A}^2 \mathcal{H}^H + \sigma^2 \mathbf{I})^+ \mathbf{H}_d}{\mathbf{H}_d^H (\mathcal{H} \mathbf{A}^2 \mathcal{H}^H + \sigma^2 \mathbf{I})^+ (\mathcal{H} \mathbf{A}^2 \mathcal{H}^H + \sigma^2 \mathbf{I})^+ \mathbf{H}_d} \\
&= \frac{\mathbf{H}_d^H \mathcal{H}^H \mathbf{A}^{-2} \mathcal{H}^+ \mathbf{H}_d \mathbf{H}_d^H \mathcal{H}^H \mathbf{A}^{-2} \mathcal{H}^+ \mathbf{H}_d}{\mathbf{H}_d^H \mathcal{H}^H \mathbf{A}^{-2} \mathcal{H}^+ \mathcal{H}^H \mathbf{A}^{-2} \mathcal{H}^+ \mathbf{H}_d} = \frac{A_d^{-4}}{\mathbf{H}_d^H \mathcal{H}^H \mathbf{A}^{-2} (\mathcal{H}^H \mathcal{H})^{-1} \mathbf{A}^{-2} \mathcal{H}^+ \mathbf{H}_d} = \frac{1}{(\mathcal{H}^H \mathcal{H})_{(d,d)}^{-1}} \quad (3)
\end{aligned}$$

where $P_d(\sigma) = Q\left(\sqrt{\frac{E_s}{E_n}}\right)$, Q is the complementary Gaussian cumulative distribution function, and “+” represents pseudoinverse. In (3) we have used the facts that $(\mathcal{H} \mathbf{A}^2 \mathcal{H}^H)^+ = (\mathcal{H}^+)^H \mathbf{A}^{-2} \mathcal{H}^+$, $(\mathcal{H}^H \mathcal{H})^{-1} = \mathcal{H}^+ (\mathcal{H}^+)^H$ and $\mathcal{H}^+ \mathbf{H}_d = [0 \dots 0 1 0 \dots 0]^H$, where 1 is in the d th position. Note the first and last equalities are based on the assumption of full column rank of channel matrix \mathcal{H} . From (3), it is seen that the AME does not depend on the interfering signal amplitudes. Thus, it is equal to the near-far resistance $\bar{\eta}_d$ [6]. It then follows that $\bar{\eta}_d = \eta_d = \frac{1}{(\mathcal{H}^H \mathcal{H})_{(d,d)}^{-1}}$. Note (3) is a generalization of the corresponding result in [6] to asynchronous multipath channels and space-time diversity scenario. \square

Proposition 1 can be carried one step further to reach an expression that facilitates comparison among various space-time settings. Before proceeding further, however, we need to define some useful matrices. Let \mathcal{I} denote the subspace spanned by interference channel vectors \mathbf{H}_i , $i \neq d$ where \mathbf{H}_i denotes the i th column in the channel matrix \mathcal{H} . $\tilde{\mathcal{H}}$ is the matrix obtained by deleting the d th column \mathbf{H}_d from \mathcal{H} . It is easy to show that $\mathcal{C}(\tilde{\mathcal{H}}) = \mathcal{I}$, where $\mathcal{C}(\cdot)$ represents the column subspace. Denote $\mathbf{M} \triangleq \mathcal{H}^H \mathcal{H}$, $\mathbf{R}_d = \tilde{\mathcal{H}}^H \tilde{\mathcal{H}}$ and $\mathbf{r}_d = \tilde{\mathcal{H}}^H \mathbf{H}_d$ is a vector resulting from deleting the d th entry from the d th column of \mathbf{M} . Note \mathbf{R}_d is non-singular due to AS3. We have the following Proposition.

Proposition 2 *The near-far resistance in (3) can be rewritten as*

$$\bar{\eta}_d = \frac{1}{(\mathcal{H}^H \mathcal{H})_{(d,d)}^{-1}} = 1 - \mathbf{r}_d^H \mathbf{R}_d^{-1} \mathbf{r}_d \quad (4)$$

Proof: (3) can be rewritten as $\bar{\eta}_d = \frac{1}{(\mathcal{H}^H \mathcal{H})_{(d,d)}^{-1}} = \frac{\det(\mathbf{M})}{(-1)^{d+d} \det(\mathbf{R}_d)}$, where $\det(\cdot)$ represents the determinant of a matrix. To compute $\det(\mathbf{M})$, let $i = d$ and do the following row and column operations on \mathbf{M} : 1) exchange the i th column with the $(i+1)$ th column; 2) exchange the i th row with the $(i+1)$ th row and set $i = i+1$. If $i \neq \text{col}(\mathcal{H})$ where $\text{col}(\mathcal{H})$ denotes the number of columns of \mathcal{H} , go to step 1; else terminate the row and column operations. We finally obtain $\bar{\mathbf{M}} = \begin{bmatrix} \mathbf{R}_d & \mathbf{r}_d \\ \mathbf{r}_d^H & 1 \end{bmatrix}$. Since from \mathbf{M} to $\bar{\mathbf{M}}$ only an even number of exchange operations are executed, $\det(\mathbf{M}) = \det(\bar{\mathbf{M}}) = \det(\mathbf{R}_d) - \mathbf{r}_d^H \mathbf{R}_d^{\text{adj}} \mathbf{r}_d$, where adj represents adjoint of a matrix and the second equality is resulted from a standard matrix equality [8] (pp. 50). Therefore, $\bar{\eta}_d = \frac{\det(\mathbf{M})}{\det(\mathbf{R}_d)} = \frac{\det(\mathbf{R}_d) - \mathbf{r}_d^H \mathbf{R}_d^{\text{adj}} \mathbf{r}_d}{\det(\mathbf{R}_d)} = 1 - \mathbf{r}_d^H \mathbf{R}_d^{-1} \mathbf{r}_d$. \square

4 Comparison of near-far resistance of MMSE detector for different space-time CDMA communication systems

It is interesting to investigate whether transmitter/receiver diversity can enhance the near-far resistance. To this end, we analyze the near-far resistance of the MMSE detector for asynchronous CDMA communication systems under frequency selective multipath channels with or without multiple transmitter and receiver antennas. Specifically, we focus on four different scenarios. 1) one transmitter antenna, one receiver antenna (uncoded); 2) one transmitter antenna, two receiver antennas (uncoded); 3) two transmitter antennas, one receiver antenna (ST coded); 4) two transmitter antennas,

two receiver antennas (ST coded)². It is assumed that the Alamouti STBC is employed in systems with two transmitter antennae.

It has been shown that the signal models of the uncoded scenarios 1 and 2 have the same form as in (2) [7]. Furthermore, the channel matrices under scenarios 1 and 2 are also of block Toeplitz structure. The near-far resistance under scenarios 1 and 2 can be derived in a similar manner as Propositions 1 and 2, and has the same expressions as (3) and (4) by substituting \mathcal{H} with the corresponding channel matrix for scenarios 1 and 2 (Actually, the results of Propositions 1 and 2 are valid for any CDMA system as long as its signal model is of the same form as (2)).

In order to facilitate fair comparison among different scenarios, we make the following assumption. **AS4**: the processing gain L_c (L_c chips per symbol), the system load J , the smoothing factor M (stack M shifted versions of the L_c -dimensional received signal vectors to form $\mathcal{X}_M(n)$ in (2)), the distributions of multipath delay spread and asynchronous user delay are the same under different scenarios. Since the dimension of the channel matrix will prove to be useful for our following derivations, we now specify those parameters. Let \mathcal{H}_i , $i = 1, \dots, 4$, denote the channel matrix corresponding to the above four scenarios. Under **AS4**, the dimensions of the channel matrix under scenarios 1 through 4 are $ML_c \times J(M + L_h^1 - 1)$, $2ML_c \times J(M + L_h^2 - 1)$, $2ML_c \times 2J(M + \lceil \frac{L_h^3}{2} \rceil)$ and $4ML_c \times 2J(M + \lceil \frac{L_h^4}{2} \rceil)$, respectively [3][7], where L_h^i (non-negative integer) is related to the maximum multipath delay spread and the maximum asynchronous user delay of the i th scenario, and is defined as in (132) and (2.11)³ of [3] and [7], respectively. Comparing the dimensions of the channel matrices of scenario 3 with scenario 1, and scenario 4 with scenario 2, it is interesting to find that with two transmitter antennas the Alamouti STBC CDMA system is equivalent to uncoded systems with virtual processing gain $2L_c$ and $2J$ virtual users.

Next we will compare the near-far resistance of MMSE detector under different scenarios. The value of the near-far resistance (3) clearly depends on the channels and asynchronous transmission delays. Since these parameters are random in nature, it is more meaningful to compare the statistical average of the near-far resistance rather than a particular random realization. To this end, we need an additional assumption. **AS5**: Under the i th scenario, assume \mathbf{H}_d^i (the vector in \mathcal{H}_i which is corresponding to the desired transmitted symbol of the desired user) is a random vector with a probability density function $\mathcal{N}_c(\mathbf{0}_{\text{row}(\mathcal{H}_i)}, \frac{1}{\text{row}(\mathcal{H}_i)} \mathbf{I}_{\text{row}(\mathcal{H}_i)})$ and is statistically independent of the interference subspace \mathcal{I} (Since it won't affect the derivation, here \mathcal{I} is a general expression which includes all 4 scenarios.), where \mathcal{N}_c represents the complex normal distribution, $\text{row}(\mathcal{H}_i)$ denotes the number of rows in \mathcal{H}_i , $\mathbf{0}_{\text{row}(\mathcal{H}_i)}$ represents the $\text{row}(\mathcal{H}_i) \times 1$ zero vector, and $\mathbf{I}_{\text{row}(\mathcal{H}_i)}$ represents the $\text{row}(\mathcal{H}_i) \times \text{row}(\mathcal{H}_i)$ identity matrix. The fact that the variance is $1/\text{row}(\mathcal{H}_i)$ is because \mathbf{H}_d^i is normalized. We have the following proposition.

Proposition 3 Denote $\bar{\eta}_d^i$, $i = 1, \dots, 4$, as the expectation of the near-far resistance of the MMSE detector under the i th scenario. Then under **AS3**, **AS4** and **AS5** for each and every above scenarios, we have 1) $\bar{\eta}_d^3 < \bar{\eta}_d^2$; 2) If $L_h^i \geq 2$, $i = 3, 4$, then $\bar{\eta}_d^1 < \bar{\eta}_d^3$, $\bar{\eta}_d^2 < \bar{\eta}_d^4$; otherwise $\bar{\eta}_d^1 \approx \bar{\eta}_d^3$, $\bar{\eta}_d^2 \approx \bar{\eta}_d^4$. Based on 1) and 2), we have $\bar{\eta}_d^1 \leq \bar{\eta}_d^3 < \bar{\eta}_d^2 \leq \bar{\eta}_d^4$.

Proof: Under **AS3**, starting from (4) and adopting the similar derivations in [6], the conditional expectation of $\bar{\eta}_d^i$, conditioning on the interference subspace \mathcal{I} , is then given by

$$\begin{aligned} E[\bar{\eta}_d^i | \mathcal{I}] &= 1 - E[\mathbf{r}_d^{iH} (\mathbf{R}_d^i)^{-1} \mathbf{r}_d^i | \mathcal{I}] = 1 - E[\text{tr}\{\mathbf{r}_d^{iH} (\mathbf{R}_d^i)^{-1} \mathbf{r}_d^i\} | \mathcal{I}] = 1 - E[\text{tr}\{\mathbf{r}_d^i \mathbf{r}_d^{iH} (\mathbf{R}_d^i)^{-1}\} | \mathcal{I}] \\ &= 1 - E[\text{tr}\{\tilde{\mathcal{H}}_d^H \mathbf{H}_d^i \mathbf{H}_d^{iH} \tilde{\mathcal{H}}_d (\mathbf{R}_d^i)^{-1}\} | \mathcal{I}] = 1 - E[\text{tr}\{\mathbf{H}_d^i \mathbf{H}_d^{iH} \tilde{\mathcal{H}}_d (\mathbf{R}_d^i)^{-1} \tilde{\mathcal{H}}_d^H\} | \mathcal{I}] \\ &= 1 - \text{tr}\{E[\mathbf{H}_d^i \mathbf{H}_d^{iH} | \mathcal{I}] \tilde{\mathcal{H}}_d (\mathbf{R}_d^i)^{-1} \tilde{\mathcal{H}}_d^H\} = 1 - \frac{1}{\text{row}(\mathcal{H}_i)} \text{tr}\{\tilde{\mathcal{H}}_d^H \tilde{\mathcal{H}}_d (\mathbf{R}_d^i)^{-1}\} \\ &= 1 - \frac{1}{\text{row}(\mathcal{H}_i)} \text{tr}\{\mathbf{R}_d^i (\mathbf{R}_d^i)^{-1}\} = 1 - \frac{\text{col}(\mathcal{H}_i) - 1}{\text{row}(\mathcal{H}_i)} \end{aligned} \quad (5)$$

²The channel matrix \mathcal{H} of scenario 4 is reduced to the channel matrix of scenario 3 when there is only one receive antenna [3].

³There is a notation error (i.e., the floor operation should be replaced by the ceiling operation) in (2.11) of [7]. Also note $\tau_{kL}^{(a,b)}$ in (132) of [3] is the sum of the corresponding multipath delay and the asynchronous user delay of user k .

where $\text{tr}(\cdot)$ represents the trace of a matrix, $\tilde{\mathcal{H}}_i$, \mathbf{r}_d^i and \mathbf{R}_d^i are defined similarly as in Section 3 for the i th scenario. The sixth equality is based on the property of conditional expectation and the seventh equality is based on the fact that $E[\mathbf{H}_d^i \mathbf{H}_d^{iH} | \mathcal{I}] = E[\mathbf{H}_d^i \mathbf{H}_d^{iH}] = \frac{1}{\text{row}(\tilde{\mathcal{H}}_i)} \mathbf{I}$ due to **AS5**. Note (5) is an extension from the similar result shown in [6] under AWGN channel case for DS-CDMA. However, in order to obtain (5) in the asynchronous multipath channel case, in **AS5** we assume a much more restrictive assumption (\mathbf{H}_d^i and vectors in \mathcal{I} are the combination of multipath channel and spreading code while in [6] the statistical independence assumption is only between different spreading codes). Based on (5) and **AS4**, it is straightforward to show that

$$E[\bar{\eta}_d^1 | \mathcal{I}] = 1 - \frac{J(M + L_h^1 - 1) - 1}{M L_c} \quad (6)$$

$$E[\bar{\eta}_d^2 | \mathcal{I}] = 1 - \frac{J(M + L_h^2 - 1) - 1}{2M L_c} \quad (7)$$

$$E[\bar{\eta}_d^3 | \mathcal{I}] = 1 - \frac{2J(M + \lceil \frac{L_h^3}{2} \rceil) - 1}{2M L_c} \quad (8)$$

$$E[\bar{\eta}_d^4 | \mathcal{I}] = 1 - \frac{2J(M + \lceil \frac{L_h^4}{2} \rceil) - 1}{4M L_c} \quad (9)$$

Under **AS4**, J , M , L_c , the maximum multipath delay spread and the maximum asynchronous user delay are the same under different scenarios for each realization of \mathcal{I} (these parameters in general may be different from one realization of \mathcal{I} to another). By carefully examining (132) and (2.11), (145) and (2.17) and the definitions of parameters in those equations in [3] and [7], respectively, it is straightforward to show that $L_h^1 = L_h^2 = L_h^3 + 1 = L_h^4 + 1$ for each realization of \mathcal{I} . Since $J(M + L_h^2 - 1) = JM + JL_h^3 < 2JM + 2J\lceil \frac{L_h^3}{2} \rceil$ for each realization of \mathcal{I} where we have used the fact that $\gamma \leq 2\lceil \frac{\gamma}{2} \rceil$ when γ is a non-negative integer, by comparing (7) and (8), we have $E[\bar{\eta}_d^3 | \mathcal{I}] < E[\bar{\eta}_d^2 | \mathcal{I}]$ for each realization of \mathcal{I} .

When $L_h^i \geq 2$, $i = 3, 4$, by comparing (6) and (8), (7) and (9), we obtain $E[\bar{\eta}_d^1 | \mathcal{I}] < E[\bar{\eta}_d^3 | \mathcal{I}]$ and $E[\bar{\eta}_d^2 | \mathcal{I}] < E[\bar{\eta}_d^4 | \mathcal{I}]$ for each realization of \mathcal{I} , where we have used the fact that $\lceil \frac{\gamma}{2} \rceil - 1/(2J) < \gamma - 1/J$ when $\gamma \geq 2$ is an integer. When $L_h^i = 0, 1$, $i = 3, 4$, $E[\bar{\eta}_d^1 | \mathcal{I}] - E[\bar{\eta}_d^3 | \mathcal{I}] = 1/(2M L_c)$ and $E[\bar{\eta}_d^2 | \mathcal{I}] - E[\bar{\eta}_d^4 | \mathcal{I}] = 1/(4M L_c)$ for each realization of \mathcal{I} . Since $1/(2M L_c)$ and $1/(4M L_c)$ are negligible for values of M and L_c in practical systems, $E[\bar{\eta}_d^1 | \mathcal{I}] \approx E[\bar{\eta}_d^3 | \mathcal{I}]$ and $E[\bar{\eta}_d^2 | \mathcal{I}] \approx E[\bar{\eta}_d^4 | \mathcal{I}]$ for each realization of \mathcal{I} .

Since $\bar{\eta}_d^i = E_{\mathcal{I}}[E[\bar{\eta}_d^i | \mathcal{I}]]$, $i = 1, \dots, 4$, the claims in Proposition 3 are proven. \square

Remarks: 1) Transmitter and receiver diversity does have benefits on near-far resistance compared with systems without those diversity (i.e., $\bar{\eta}_d^1 \leq \bar{\eta}_d^3$, $\bar{\eta}_d^1 < \bar{\eta}_d^2$). Furthermore, receiver diversity has more impact on the near-far resistance (i.e., $\bar{\eta}_d^3 < \bar{\eta}_d^2$). 2) Although Propositions 1, 2 and 3 are derived based on Alamouti STBC for up to two-transmit/receive antenna configurations, these results can be extended to any STBC for any transmit/receive antenna configurations as long as the signal model has the same form as in (2).

Finally, it is worth to point out that **AS3** is not indispensable to obtain the near-far resistance of the MMSE detector for Alamouti ST coded CDMA system. However, Due to lack of space, detailed derivations of corresponding results and how spatial diversity may affect the rank of the channel matrix will be presented in another paper.

5 Simulations

In this section, simulations are conducted to verify the theoretical findings. Gold sequence of length L_c is employed as user spreading sequences. The multipath channels for each user and each transmit-receive pair have N_p paths with a total delay spread (including the user transmission delay) of two symbol durations, and all multipaths have mutually independent delays uniformly distributed over two

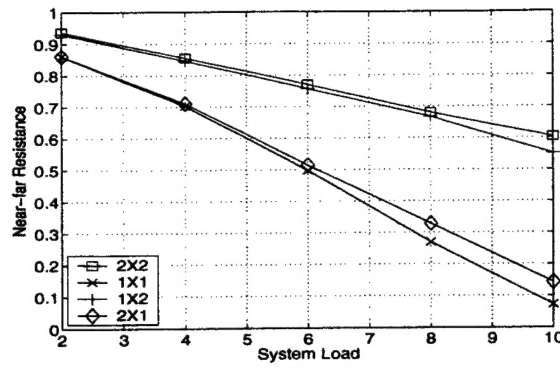


Figure 5.1: Near-far Resistance Comparison Among Different transmit/receive antenna Settings.

symbol intervals. All N_p multipath amplitudes are mutually independent, complex Gaussian with zero-mean and unit variance. The spreading sequences and multipath channels for each user are randomly generated in each of the Monte Carlo run (i.e., they are different in different runs).

Example: Near-far resistance comparison among different antenna configurations

In this simulation, we compare the theoretical near-far resistance of MMSE detector among different space-time settings. $L_c = 15$ and $N_p = 6$. The smoothing factor $M = 3$. All near-far resistances are calculated by averaging 2000 Monte Carlo runs. The result is presented in Fig. 1. It can be seen from Fig. 1 that the simulation result has a good agreement with the theoretical findings in Proposition 3.

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